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FASTC-ID(RS)T-0309-92

FOREIGN AEROSPACE SCIENCE AND TECHNOLOGY CENTER

DETECTION PERFORMANCE OF DIGITAL POLARITY
SAMPLED PHASE REVERSAL CODED PULSE COMPRESSORS

by

Zhu Zhaoda, Tu Shude

HUMAN TRANSLATION

FASTC-ID(RS)T-0309-92 2 April 1993

MICROFICHE NR: 93C000242

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English pages: 8

Source: Dianzikexue Xuekan, Vol. 9, Nr. 3, May 1987; pp. 269-272

Country of origin: China
Translated by: SCITRAN
F33657-84-D-0165
Requester: FASTC/TAER/Lt Mike Sutton

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FASTC - ID(RS) T - 0309 - 92

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Zhu Zhaoda and Tu Shude

(Nanjing Aeronautical Institute)

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May, 1987

Translated by:

S C I T R A N

1482 East Valley Road

Santa Barbara, CA 93150

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(Received Dec. 25, 1985. Accepted Aug. 1, 1986)

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Abstract

The nonparametric constant false alarm rate (CFAR) property of digital polarity sampled phase reversal coded pulse compressors is described. The detection performance in Gaussian and non-Gaussian noise is determined. It is shown that the loss in signal-to-noise ratio of the processor, relative to the incoherent matched filter, decreases as the code length increases, the asymptotic loss in Gaussian noise is 1.96dB, and the loss in Weibull noise decreases with the shape parameter of the Weibull distribution and can even become a gain.

I. Introduction

While doing pulse compressing, digital polarity sampled phase reversal coded pulse compressors[1] also completes constant false alarm rate (CFAR) processing. This type of processors are easy to implement and thus are widely used in radars. Ref [2] has analyzed the loss mechanism of this type of processors and given the quantitative study of the loss in typical applications with Gaussian noise and finite code length. This paper studies the non-parametric CFAR characteristics and the detection performance in Gaussian and non-Gaussian noise of this type of processors. We derive the asymptotic relative efficiency (ARE) in Gaussian noise, and determine the loss of signal to noise ratio by using analytic and numerical calculations. Part of the material has been discussed in Ref [3].

II. Non-parametric CFAR characteristics

The schematic diagram of the processors is shown in Fig. 1. When the signal is present, the complex variable representation of the mid-frequency received wave form is

$$z_i = V_i\, e^{j\phi_i} = A\, e^{j\phi_i} + N_i\, e^{j\phi_i},\quad i=1,\ \cdots,\ M, \tag{1}$$

where V_i, A and N_i represent received wave form, signal and noise amplitude corresponding to the i-th code element; ϕ_i, θ_i, and ψ_i indicate their corresponding phases. Among them the phase of the signal $\theta_i=\theta_0+\alpha_i$, and θ_0 is the initial phase. Depending on the coding, α_i takes values of either 0 or π. M is the code length. The synchronous phase of z_i and the orthogonal components of polarity samples $\mathrm{sgncos}\phi_i$ and $\mathrm{sgnsin}\phi_i$ are separately transferred to the duel signal registers and are compared digit by digit with the codes stored in the coding registers. If they match, +1 is chosen. Otherwise -1 is chosen. Sum up the results of the comparison to get the total of each circuit branches

$$I = \sum_{i=1}^{M} \mathrm{sgncos}\phi_i\ \mathrm{sgncos}\alpha_i, \tag{2}$$

$$Q = \sum_{i=1}^{M} \mathrm{sgncos}\phi_i\ \mathrm{sgncos}\alpha_i, \tag{3}$$

The detection statistics

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$$t = \sqrt{I^2 + Q^2} \tag{4}$$

is obtained.

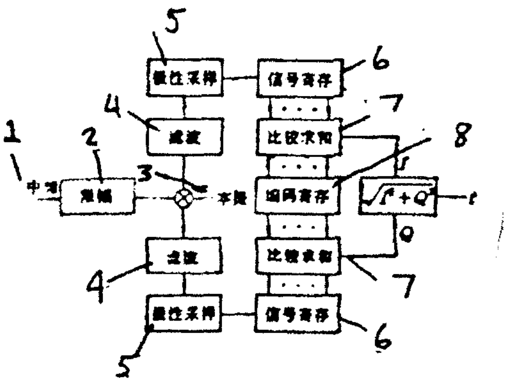

Fig. 1 Processor schematic diagram. 1. middle frequency signal, 2. amplitude limiter, 3. self excitation, 4. filtering, 5. polarity sampling, 6. signal register, 7. comparison adder, 8. coding register.

When the noise phase ψ_i independently co-distributes and the distribution of ψ_i does not depend on the noise amplitude distribution, the false alarm probability of this type of processors is independent of the noise amplitude distribution, i.e. it has CFAR characteristics. In radar applications, circular symmetric noise distribution is often encountered. Its amplitude and phase are mutually independent and the phase distribute uniformly on $(0, 2\pi)$. The sampling circuit generally uses code sampling and the samples are independent. Under these circumstances, the processors have the non-parametric CFAR capability. This type of processors can also be regarded as a simplified version of Dicke-fix receivers[4], or a symbol detector of narrow band signals[5].

III. Asymptotic performances in Gaussian noise

Due to the influence of the noise, ϕ_i in general is different from θ_i. In Gaussian noise and non-undulating signal situation, the Pete error probability of the circuit branch I, i.e. the probability that $\cos\phi_i$ and $\cos\theta_i$ has opposite signs is[6]

$$q_1 = 1 - \Phi(\sqrt{2S}\ |\cos\theta_0|), \tag{5}$$

where

$$\Phi(x) = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{x} \exp(-\frac{u^2}{2})\ du$$

is the Gaussian distribution function, S the input signal-to-noise ratio for the processors shown in Fig. 1. Similarly the Pete error probability of the circuit branch Q is

$$q_2 = 1 - \Phi(\sqrt{2S}\ |\sin\theta_0|), \tag{6}$$

Since α_i is either 0 or π, when $-\frac{\pi}{2} \le \theta_0 < \frac{\pi}{2}$ and $\cos\theta_i$ and $\cos\alpha_i$ take the same sign, the probability of $\mathrm{sgn}\cos\phi_i \mathrm{sgn}\cos\alpha_i$ equaling to +1 is $1-q_1$, and the -1 probability is q_1. Hence I obeys binomial distribution, and its probability is

$$P_I(n = -M + 2N) = \binom{M}{N} (1-q_1)^N q_1^{M-N}, \ N = 0, 1, \cdots, M. \tag{7}$$

When $\frac{\pi}{2} \le \theta_0 < \frac{3\pi}{2}$ and $\cos\theta_i$ and $\cos\alpha_i$ take opposite signs, the probability of I is the same as formula (7) except $n = M - 2N$. Only I^2 will be used in the following, hence formula (7) can always be used for any value of θ_0. Similarly, the probability of Q is

$$P_Q(n = -M + 2N) = \binom{M}{N} (1-q_2)^N q_2^{M-N}, \quad N = 0,1, \cdots, M. \qquad (8)$$

Since the two orthogonal components of the narrow band Gaussian noise are independent, variables I and Q are independent.

For very large M and very small S, t converges to Rice distribution. Its probability density function is

$$p(t) = \frac{t}{M} \exp\left(-\frac{t^2+4M^2S/\pi}{2M}\right) I_0(2t\sqrt{S/\pi}\,), \quad t \geq 0. \qquad (9)$$

When only noise is present, S=0, formula (9) becomes Rayleigh distribution. Its probability density function is

$$p_0(t) = \frac{t}{M} \exp\left(-\frac{t^2}{2M}\right), \quad t \geq 0. \qquad (10)$$

The relationship between the detection probability P_d and the false alarm probability P_f is

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$$P_d = Q\left(\sqrt{2\ln\frac{1}{P_f}}, \sqrt{\frac{4M\ S}{\pi}}\right). \qquad (11)$$

Here $Q(\cdot)$ denotes Marcum Q function.

For the Gaussian noise, the best parameter detector for detecting phase coded signals with unknown initial phase is the non-interfering matching filter. When the input signal to noise ratio is S_L and the code length is M_L, the detection probability P_d and the false alarm probability P_f of the non-interfering matching filter satisfy the following relation[7]:

$$P_d = Q\left(\sqrt{2\ln\frac{1}{P_f}}, \sqrt{2M_LS_L}\right). \qquad (12)$$

In order to make the P_d and P_f of the two kinds of detectors equal, the following has to be satisfied

$$\frac{4M\ S}{\pi} = 2\ M_LS_L. \qquad (13)$$

Hence, the asymptotic relative efficiency of the processor shown in Fig. 1 compared with the non-interfering matching filter is

$$ARE = \lim_{S=S_L\to 0} \frac{M_L(P_d, P_f, S_L)}{M(P_d, P_f, S_L)} = \frac{2}{\pi}. \qquad (14)$$

Similarly, the asymptotic loss is

$$L_\infty = \lim_{M=M_L\to\infty} \frac{S(P_d, P_f, S_L)}{S_L(P_d, P_f, S_L)} = \frac{\pi}{2},$$

$$L_\infty(\mathrm{dB}) = 1.96. \tag{15}$$

For undulating signals, (11) and (12) should be averaged according to the undulating model. Then formula (13) will still be valid to the averaged signal to noise ratio. Therefore ARE and L_∞ do not change.

IV. Finite code length performance

The probability distribution of I^2 and Q^2 are respectively

$$P_{I^2}(n = m) = P_I(n=\sqrt{m}) + P_I(n=-\sqrt{M}),\ m>0,$$
$$P_{I^2}(n = 0) = P_I(n=0) \tag{16}$$

and

$$P_{Q^2}(n = m) = P_Q(n=\sqrt{m}) + P_Q(n=-\sqrt{M}),\ m>0,$$
$$P_{Q^2}(n = 0) = P_Q(n=0). \tag{17}$$

Under Gaussian noises, I^2 and Q^2 are independent variables. Therefore, the probability of t^2 is

$$P_{t^2}(n) = P_{I^2}(n) * P_{Q^2}(n), \tag{18}$$

where * denotes convolution. From the noise only $P_{t^2}(n)$ we can determine the threshold corresponding to the specified false alarm probability P_f. Afterwards, based on this threshold and the value of $P_{t^2}(n)$ for both the signal and noise, we can obtain the detection probability P_d for non-undulating signal. Note that P_d is related to θ_0. The detection performance of non-interfering matching filter can be obtained from Ref [7]. The calculation result is summarized in Fig. 2. Shown in the figure is the relationship between the signal to noise ratio loss and M for the processor of Fig. 1 relative to that of non-interfering matching filter under non-undulating signal and when θ_0=45°, P_d=0.5, $P_f=10^{-4}$ and 10^{-6}.

For non-Gaussian noise, the calculation assumes Weibull noise model. Since the processor in Fig. 1 has the non-parametric CFAR capability, the threshold corresponding to a given P_f value in Weibull noises is the same as that determined under Gaussian noises in the above. When calculating P_d, Monte Carlo simulation is used due to lack of analytic solutions. The detector performance of the non-interfering matching filter under Weibull noises is also obtained by Monte Carlo simulation. The calculation result is summarized in Fig. 3. Shown in the figure is the relationship between the signal to noise ratio loss and the Weibull shape factor α for the processor of Fig. 1 relative to that of non-interfering matching filter under non-undulating signal and when $\theta_0=45°$, $P_d=0.5$, $P_f=10^{-4}$ and 10^{-6}. The tail of the Weibull noise envelop distribution becomes longer as α decreases. The $\alpha=2$ Weibull distribution is the narrow band Gaussian noise situation. We can see that the loss of signal to noise ratio decreases as α decreases. It even becomes a gain.

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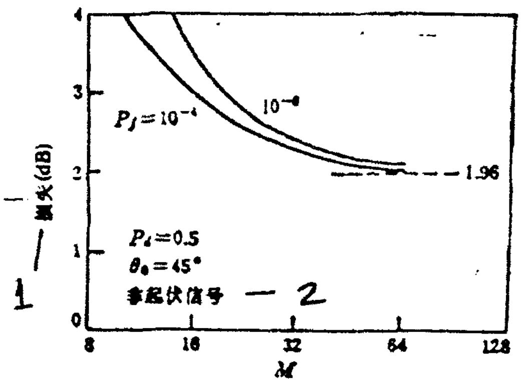

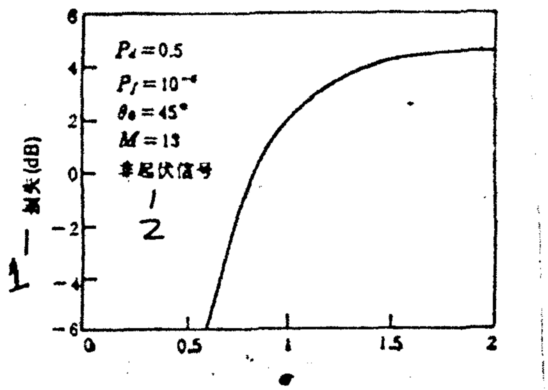

Fig. 2 Loss in Gaussian noise. Fig. 3 Loss in Weibull noise.
1. loss, 2. non-undulating signal.

V. Conclusion

Polarity sampled phase reversal coded pulse compressors have the nonparametric CFAR capability and are easily realizable. Compared to non-interfering matching filters, the relative detection performance of this kind of processors improves as the code length and the tail of the noise envelop distribution increase. In Gaussian noises, their asymptotic loss is 1.96dB.

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